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A Comparison of Multi-Carrier OFDM and Single Carrier Iterative Equalisation for Future High Performance Wireless Local Area Networks

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Abstract- There has been a strong trend towards the specification of multi-carrier Coded Orthogonal Frequency Division Multiplexing (COFDM) in the physical layer of emerging high performance Wireless Local Area Networks (WLAN). This trend is based on an alleged combination of enhanced performance and lower terminal complexity. However, in recent years significant advances have been made in the field of single carrier equalisation, where iterative equalisation and decoding now offers excellent error rate performance in slow and fast fading channels.

This paper investigates the possibility of employing a single carrier system using iterative equalisation at the receiver as a direct competitor to multi-carrier COFDM. The throughput, range and terminal complexities of a COFDM solution are compared with those of a single carrier iterative equalised approach for identical indoor channel conditions. The paper demonstrates that iterative equalisation outperforms COFDM in an indoor environment. However, from a cost/performance viewpoint, COFDM is still seen as the more attractive solution.

Keywords- Iterative Equalisation, Orthogonal Frequency Division Multiplexing, Turbo Coding

I. INTRODUCTION

There has been a strong trend towards the specification of multi-carrier Coded Orthogonal Frequency Division Multiplexing (COFDM) in the physical layer of emerging high performance wireless local area networks (WLAN). ETSI Hiperlan /1 was the last high speed wireless LAN standard to specify the use of single carrier adaptive equalisation. The trend towards COFDM is based on an alleged combination of enhanced performance and lower terminal complexity. However, in recent years significant advances have been made in the area of single carrier equalisation. This paper aims to compare the throughput, range and terminal complexity of a COFDM solution with that

of a single carrier iterative equalised approach over identical indoor channel conditions.

The IEEE 802.11 family is a good example of the latest COFDM standards, where 802.11a uses a 64 sub-carrier COFDM approach to support data rates up to 54 Mbits/s in favourable operating conditions [1]. More recently, 802.11g has specified the use of COFDM in the 2.4 GHz band. By employing a number of orthogonal narrowband sub-carriers, COFDM mitigates the harmful effects of inter-symbol interference (ISI) while utilising forward error correction to benefit from the frequency diversity present in a wideband channel. COFDM can tolerate large values of delay spread by using a guard interval and a single tap per sub-band frequency domain equaliser. However, the use of a direct inversion frequency domain equaliser prevents the use of the latest turbo iteration concepts.

Since the initial proposal of Turbo Codes by Berrou et al in 1993 [2], the iterative principle has been extended to encompass single carrier equalisation techniques [3]. This allows single carrier systems to combine the operations of channel coding and equalisation to operate in a wideband channel with performances that could not previously be achieved with traditional equalisation and forward error correcting (FEC) techniques. Iterative equalisation techniques have been shown to give excellent error rate performance in slow and fast fading channels [4].

This paper investigates the possibility of employing a single carrier system using iterative equalisation at the receiver as a direct competitor to a multi-carrier COFDM approach. The paper aims to determine which technique offers the highest academic performance, regardless of terminal complexity. A baseband software simulation (broadly based on the 802.11a parameters) of the proposed COFDM and single carrier iterative equaliser schemes has been generated. In order to implement a fair system comparison, the packet structure of both systems is based on a data rate of 20 Mbits/s and a common convolutional encoder and interleaver. Identical mean power transmit levels and antenna gains are also assumed for data rate versus range predictions.

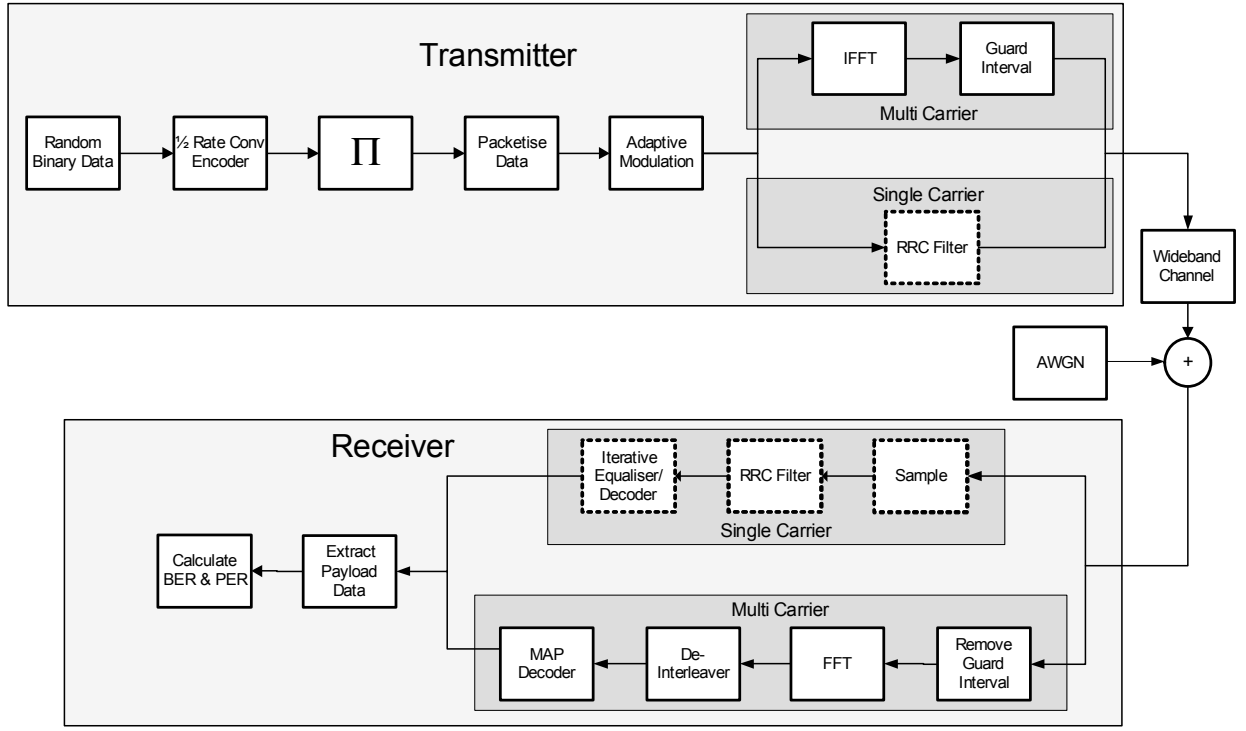


Figure 1. Baseband Simulation Block Diagram

II. SIMULATION STRUCTURE

For the purpose of this study, the baseband simulation shown in Figure 1 has been developed.

A. Encoding

The payload element of a packet is made up of 574 data bits, which are encoded with the half rate recursive systematic convolutional (RSC) encoder shown in Figure 2, to form a data block of 1152 bits. The final 4 bits are appended to ensure that the encoder is returned to the all zero state at the end of the block. This is achieved by switching off the input bit stream and considering only the feedback bits as the input to the encoder [5]. The RSC encoder can be described by the generator polynomial given in (1). This encoder is of input constraint length $CL=2$ and therefore produces a trellis structure with $2^{CL} = 4$ states.

$$G = \begin{bmatrix} 1, \frac{101}{111} \end{bmatrix} \begin{matrix} \text{Output} \\ \text{Feedback} \end{matrix} \quad (1)$$

B. Interleaving

The encoded data block is then passed through a random interleaver of size 1152 elements, represented in figure 1 by the Π symbol. This is necessary to allow iterative equalisation at the receiver and improves the error correcting capability in the multi-carrier system. The interleaver ensures de-correlation between the data block and the symbols that are passed through the channel.

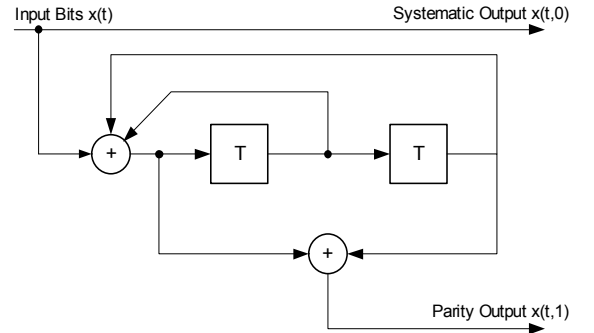


Figure 2. Recursive Systematic Encoder

C. Packet Structure

The payload is formed into a packet following the structure shown in Figure 3. A 128 bit training sequence is attached to the start of the packet, to allow for channel estimation with the LMS algorithm [6] in the single carrier system and with direct computation in the frequency domain for the multi-carrier system. Simulation has shown that this is of adequate size to allow LMS convergence for the channel used in this study. Also included is a 64-bit header which can be used for information such as the modulation mode in the following payload block.



Figure 3. Packet Structure

Once formed, the packet is modulated, using (in this study) BPSK modulation for the entire packet. A scheme using adaptive modulation, depending upon the prevailing channel conditions could be employed to maximise the user data rate.

D. Multi-Carrier Transmission

The modulated symbols before transmission are combined in a 64 symbol OFDM packet structure and passed through a 64-point IFFT. A guard interval is then attached at the beginning of each OFDM symbol. Each guard is the repetition of the last 16 samples (800 ns) of each OFDM symbol. This cyclic repetition protects the data from intercarrier interference (ICI) in a time dispersive channel.

E. Single-Carrier Transmission

Prior to transmission, the modulated data is filtered with a root raised cosine (RRC) filter with a roll-off factor $\alpha = 0.4$ and a span of 10 symbols. Up-sampling is performed in the filter to a rate of 5 samples per symbol; hence the filter comprises 50 taps.

F. Channel Model

ETSI channel model A is used in this study. This model was originally developed to evaluate the Hiperlan /2 (and later the 802.11a) standard and represents an average indoor power delay profile for a small office environment for transmissions in the 5.2GHz band [7]. The channel model has an RMS delay spread of 50ns and a maximum delay of 390ns, corresponding to a maximum symbol span of 8 symbols in the single carrier system. Each of the taps is assumed to suffer from independent Rayleigh fading statistics.

G. Multi-Carrier Receiver

Initially, the cyclic repetition is removed from the received data. The resulting vector is then passed through a 64-point FFT. Channel State Information (CSI) is calculated by comparing the original training sequence to the received symbols. The CSI is then used to compute the required taps for a single tap per carrier Frequency Domain Equaliser.

The equalised data is then de-interleaved and decoded using a SISO MAP decoder identical to the constituent decoder present in the iterative equaliser.

H. Iterative Equalisation

Figure 4, shows the iterative equalisation receiver structure used in this study [4]. Both the equaliser and the decoder employ the optimal symbol by symbol Maximum A-Posteriori (MAP) soft input soft output (SISO) algorithm [8]. Soft input symbols are fed into the decoder from a sampled receive filter stream $r(t)$ and bit-wise hard decisions are produced as the final output.

The RSC encoder and the channel can be considered as a serially concatenated coding scheme, similar to that of a serially concatenated turbo encoder [9]. This observation means that it is possible to replace the first MAP decoder in a serial turbo receiver with a MAP equaliser. It is possible to equalise and decode in an iterative manner that is similar to turbo decoding.

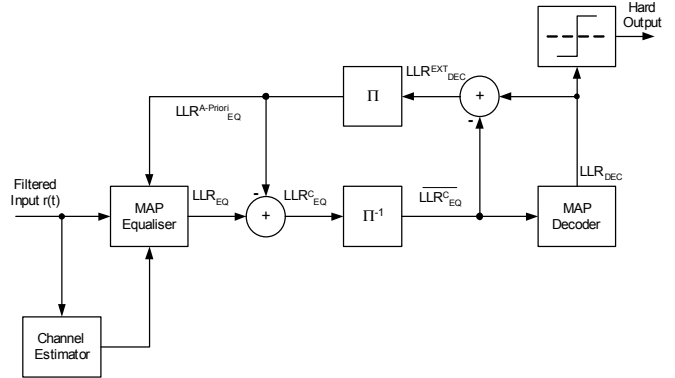


Figure 4. Iterative Equaliser Structure

III. PER RESULTS AND DISCUSSION

Figure 5, shows Packet Error Rate (PER) versus average SNR results for the multi-carrier OFDM and single carrier iterative equalisation schemes (up to three iterations).

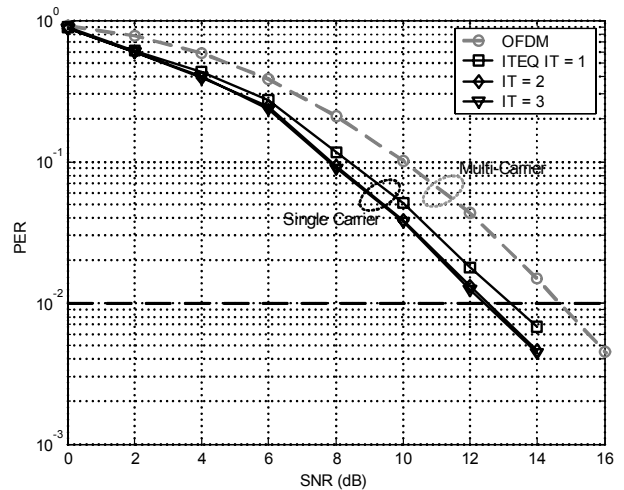


Figure 5. Packet Error Rate Comparison (ETSI channel A)

Three major trends can be observed from figure 5. At a target packet error rate of 1%, which is acceptable for most communication systems, the single carrier system has a gain of 2.3dB on the third iteration over the multi-carrier system. At the same target error rate, there is a gain of 0.7dB from the first to the second iteration for the single carrier system and negligible gain from the second to the third iteration. At lower SNRs, the iterative gain diminishes and is negligible at 4dB SNR.

IV. DATA RATE ANALYSIS

In order to quantify the PER results, a data rate and range analysis is performed as a function of SNR using a simple link budget. The instantaneous data rate DR versus SNR_{dB} can be calculated from the PER as follows:

$$DR_{SNR} = (R_S \times K) \times \left(\frac{k}{n}\right) \times (1 - PER_{SNR}) \quad (2)$$

where K represents the number of bits per symbol, k the number of information bits and n the total number of bits per packet, i.e. (k/n) is the packet code rate. It is assumed that no packets are discarded due to header errors. The PER, and therefore the DR, is purely a function of the payload error rate. Table 1, gives the rate efficiency, defined as the user data rate over the symbol rate of each system assuming the packet structure already presented. The rates are low due to the presence of the $\frac{1}{2}$ rate encoder and training/header overheads.

System	Rate Efficiency	Max. User Rate
Multi-Carrier	34.2%	6.83 Mbits/s
Single-Carrier	42.7%	8.54 Mbits/s

Table 1. Rate Efficiency

$$SNR_{dB} = P_R - N \quad (3)$$

In order to examine the achievable data rates versus range, a link budget must be established. To calculate the SNR at the receiver, given by (3), we must consider expressions for the noise power N and also the receiver power P_R in terms of the transmit power P_T and the path loss P_L :

$$N = 10 \log_{10}(K_B T) + 10 \log_{10}(R_S \times (1 + \alpha)) + NF \quad (4)$$

where $K_B = 1.38e-20$ mW/Hz/Kelvin, $T=300$ Kelvin and NF is the receiver noise figure and:

$$P_R = P_T - P_L$$

$$P_L = G_T + G_R + 20 \log_{10}\left(\frac{4\pi d}{\lambda}\right) + Fl + AF * d \quad (5)$$

where Fl is a fixed path loss associated with an obstruction to the line-of-sight path (resulting in Rayleigh fading as assumed in ETSI channel A) and AF is an attenuation factor caused by clutter in the environment. The transmission parameters described in table 2, are assumed in the link budget.

Parameter	Value
R_S	20 MHz
α_{mc}	0
α_{sc}	0.4
N_F	10 dB
P_T	0 dBm
G_T	0 dB
G_R	0 dB
Fl	6 dB
AF	0.2, 0.5, 1.0 dB/m
f_c	5.2 GHz
λ	0.057 m

Table 2. Transmission Parameters

Figure 6 shows the instantaneous achievable data rate versus average SNR for both the multi carrier and the third iteration of the single carrier system. For a nominal target data rate of 5Mbits/s, the single carrier system has a gain of 3.4dB over the multi-carrier system. The maximum data rates achieved at 14dB SNR in both systems are approximately 8.5Mbits/s and 6.8Mbits/s for the single and multi-carrier systems respectively. For all SNR values, the single carrier system achieves a greater instantaneous data rate.

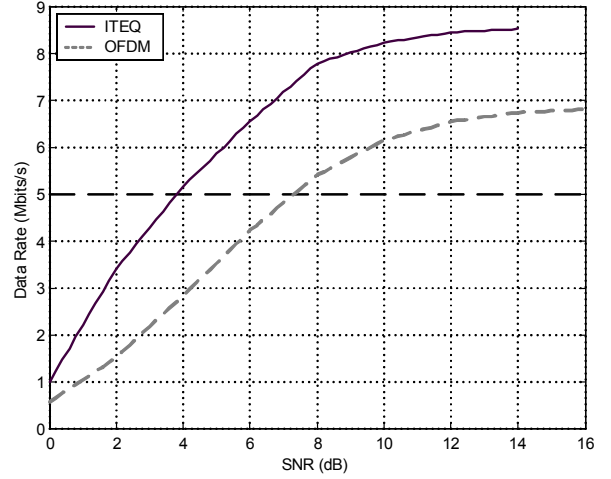


Figure 6. Date Rate vs. SNR Comparison

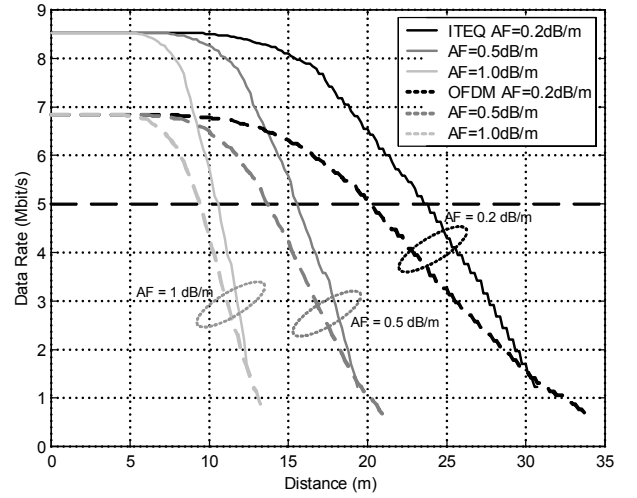


Figure 7. Date Rate vs. Range Comparison

Figure 7, shows data rates for both systems versus range from the transmitter to the receiver for varying values of attenuation factor (AF). This gives an indication of the achievable data rate for a number of non line-of-sight operation environments with varying attenuation factor (AF), representing varying amounts of clutter. As the amount of clutter increases, corresponding to a greater value of AF , then the range at which a nominal data rate of 5Mbits/s can be achieved falls for both systems. In the highly cluttered environment ($AF=1$ dB/m), then the target rate can only be

achieved at a range of 9m and 11m for the multi and single carrier systems respectively. For less cluttered environments, the range increases, as does the range advantage obtained by the single carrier system. For $AF=1\text{dB/m}$, the single carrier range advantage is only 2m, rising to 4m for the least cluttered environment ($AF=0.2\text{dB/m}$).

V. CONCLUSIONS

Results presented here show that a single carrier system employing iterative equalisation can outperform a multi-carrier system using COFDM in terms of PER and data rate. The operating bandwidth of the iterative solution is however greater than the multi-carrier system. This is due to the use of a common symbol rate in the comparison and the inclusion of a RRC filter in the transmit chain of the single carrier system. The filter adds redundancy in the frequency domain and is necessary to ensure the suppression of out-of-band frequency components that would otherwise result in interference to neighbouring spectrum users. In the multi-carrier case, redundancy is added in the time domain in the form of a cyclic repetition to combat the effect of ICI. This manifests as a reduction in the maximum available user data rate. The guard interval used here can mitigate ICI to a maximum excess delay of 800ns. It is also common for a number of null sub-bands to be specified in COFDM to ease spectral filtering. Severe frequency domain filters can then be used with the resulting group delay suppressed using the guard interval. Channel model A has a maximum delay spread of 390ns and ignoring RF filter group delay, the COFDM guard interval could be reduced, resulting in a maximum user data rate of 7.6Mbits/s.

There is a clearly a trade off between the performance of the iterative equalisation, with its associated receiver complexity, and the inherently simple OFDM receiver architecture. The complexity of the iterative equaliser is dominated by the complexity of the MAP equaliser. The number of states in the equaliser trellis is dependent upon both the modulation order and the memory of the channel. In a MAP decoder the trellis size is only dependent on the constraint length L of the code.

There are a number of different approaches to describing the complexity of the MAP algorithm [10,11,12]. Using the technique described in [12] the relative complexity of the MAP equaliser for BPSK, QPSK and 16-QAM modulation modes and channel model A with 8 symbol spaced taps, relative to the complexity of a MAP decoder operating on a $\frac{1}{2}$ rate code of constraint length 2, is given in table 3. The complexity calculations assume that the algorithms are full complexity. The calculations are per output bit and take no account of memory requirements.

Modulation Order	Equaliser Complexity
2	64
4	36981
16	20830591386

Table 3, MAP Equalisation Complexity Relative to a $\frac{1}{2}$ Rate MAP Decoder

When receiver complexity is considered we conclude that it is infeasible to use this type of iterative equalisation technique with high modulation orders combined with channels with large delay spreads due to the huge relative complexity. It should be noted that the receiver complexity of COFDM based systems is independent of the modulation order and also of channel length. Hence, although for BPSK the use of single carrier iterative equalisation is attractive in short delay spread indoor channels, this would not be the case for higher level modulation schemes or channels with greater symbol memory.

Active research to reduce the complexity of iterative equalisation is on-going, and a recent filter based technique has been reported whose complexity is independent of modulation order [13]. This approach would be particularly attractive for higher order single carrier modulation schemes. Future work will compare the complexity of this new method to that of the MAP based approach presented here.

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